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Method for equalizing a received signal

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Technical field

The invention relates to a method for equalizing a received signal in a digital receiver with the aid of a DFE (Decision Feedback Equalizer) structure, the received signal being based on a signal constellation which is one-dimensional or can be transformed to be one-dimensional.

Prior art

The transmition channels typically occurring in the case of GSM (Global System for Mobile Communication), HIPERLAN (High PErformance Radio Local Area Netzwork), DECT (Data Exchange for Cordless Telephone) etc. are characterized by the interfering effects of multipath propagation.

It is known that a Decision Feedback Equalizer (DFE) can be used in order to equalize in the digital communication system a signal which has been disturbed by a linear frequency-selected process (such as the multipath propagation in a radio channel, for example).

The performance of a DFE depends on the quality with which the filter coefficients are calculated and/or fixed in the feedforward part and in the feedback part. In the case of an unknown channel, the coefficients are typically fixed by adaptive training. If the pulse response of the channel is known, by contrast, the optimum coefficients of the DFE can then be derived therefrom.

The structure of a DFE is very simple per se and therefore very readily used. However, it is not always possible to achieve the desired performance.

Summary of the invention

The object of the invention is to specify a method of the type mentioned at the beginning which permits the determination of optimum coefficients with as little outlay on computation as possible on the

basis of the known and/or previously estimated channel unit pulse response, an enhanced performance being achieved at the same time by comparison with the known DFE in accordance with the prior art.

The features of Claim 1 define the achievement of the object. In accordance with the invention, the coefficient of the DFE are fixed so as to minimize the expected value of the squared real part of the error.

By contrast with the prior art, the error, which is a complex value per se, is not used as a basis for optimization. However, calculation is limited to the real value. The filter coefficients of the feedback filter are not complex, only those of the feedforward filter being so in general. The essential point is that the performance of the DFE structure can be improved in this basically simple way, it even being possible to reduce the computational outlay in comparison with the prior art.

In the case of a binary BPSK signal the coefficients are preferably calculated in accordance with the formulas (I) and (II) specified further below.

The invention is suitable not only for BPSK (BPSI [sic] = Binary Phase Shift Keying) signals, but also for GMSK and OQPSK modulation methods (GMSK = Gaussian Minimum Shift Keying, OQPSK = Offset Quadrature Phase Shift Keying). Also to be regarded as one-dimensional modulation methods are, therefore, those which although having a two-dimensional signal constellation can be transformed (with the aid of a suitable transformation) into an (at least approximately) equivalent one-dimensional representation.

The circuitry for implementing the method according to the invention poses no special difficulties. The calculation is typically programmed in a processor or ASIC.

The invention is suitable, for example, for a HIPERLAN system. (Such an advantageous system structure follows, for example, from EP 0 795 976 A2, Ascom Tech

AG). The so-called European Telecommunications Standard (ETS) defines the technical characteristics of a wireless local high performance network (HIPERLAN). HIPERLAN is a short range communications subsystem with a high data rate (compare in this regard ETSI 1995, ETS 300 652, UDC: 621 396). The ETS-HIPERLAN standard is provided for the frequency band 5.15 to 5.30 GHz.

Further advantageous embodiments and combinations of features of the invention result from the following detailed description and the totality of the patent claims.

Brief description of the drawings

In the drawings used to explain the exemplary embodiment:

Figure 1	shows a diagrammatic illustration of a DFE;
Figure 2	shows a diagrammatic illustration of an exemplary embodiment;
Figure 3	shows a representation of the performance of the method according to the invention by comparison with the prior art;
Figures 4a-c	show a comparison of the error behavior in the prior art and with the invention;
Figure 5	shows a diagrammatic illustration of a BPSK receiver;
Figure 6	shows a diagrammatic illustration of a GMSK receiver.

Ways of implementing the invention

The principle of the invention is to be stated below by a comparison with the prior art.

Figure 1 shows a block structure, known per se, of a DFE. The received signal I downwardly modulated by the carrier is entered into a feedforward filter FF of the DFE. Thereafter, it is combined (adder) with the estimated signal $\hat{I}$ fed back by the decision devide DD via the feedback filter FB. The signal $\tilde{I}$ is therefore

present at the input of the decision device DD. The coefficients **f** and **g** (which are understood in the present as vectors with a plurality of coefficient components) are calculated as follows in accordance with the prior art:

$$\min_{f,g} E\left\{\left|\tilde{I} - \hat{I}\right|^2\right\} \qquad \text{(A)}$$

In contrast therewith, the invention carries out the following optimization:

$$\min_{f,g} E\left\{\left(\mathrm{Re}\left(\tilde{I} - \hat{I}\right)\right)^2\right\} \qquad \text{(B)}$$

The difference from the prior art therefore consists in the type of calculation of the filter coefficients. The remaining structure of the DFE is maintained without change. This is explained in detail below with the aid of exemplary embodiments.

Figure 2 shows a concrete example of a DFE. As is usual for modern coherent digital receivers, the signal processed by it is represented by complex numbers. The real part stands in this case for the in phase component, and the imaginary part stands for the quadrature component. In accordance with the generally current understanding, the DFE shown in Figure 2 has complex coefficients and complex data.

If only the real part of the error is optimized according to the MMSE (MMSE = Minimum Mean Square Error) criterion, the feedforward filter coefficients are given by the following system of equations:

$$\text{(I)} \qquad h^R_{M+1-i} = \frac{\sigma^2}{2} f_i^R + \sum_{m=1}^{M} f_m^R \sum_{n=1}^{M} h^R_{n+1-i} h^R_{n+1-m} - \sum_{m=1}^{M} f_m^I \sum_{n=1}^{M} h^R_{n+1-i} h^I_{n+1-m}$$

$$-h^I_{M+1-i} = \frac{\sigma^2}{2} f_i^I - \sum_{m=1}^{M} f_m^R \sum_{n=1}^{M} h^I_{n+1-i} h^R_{n+1-m} + \sum_{m=1}^{M} f_m^I \sum_{n=1}^{M} h^I_{n+1-i} h^I_{n+1-m}$$

These are 2M real equations ($1 \leq i \leq M$). Coefficients whose indices are too great or too small are to be taken as 0 in this case. The indices run from 1 to L for vectors of length L. The values of the filter coefficients can be obtained using methods known per se for solving systems of linear equations. There is no need to go into these standard methods in more detail.

The feedback filter coefficients are determined by the following equations:

$$\text{(II)} \qquad g^R_{i-M} = -\sum_{m=1}^{M} f^R_m h^R_{i+1-m} - f^I_m h^I_{i+1-m}$$

These are N-1 equations, because $M + 1 \leq i \leq M + N - 1$.

Formulae (I) and (II) are based on the following conventions:

N	length of the channel unit pulse response;
M	length of the feedforward filter;
h_1^R	real part of the channel unit pulse response, $1 \leq i \leq N$,
h_1^I	imaginary part of the channel unit pulse response, $1 \leq i \leq N$,
f_1^R	real part of the filter coefficients of the feedforward part of the DFE, $1 \leq i \leq M$,
f_1^I	imaginary part of the filter coefficients of the feedforward part of the DFE, $1 \leq i \leq M$,
g_1^R	real part of the filter coefficients of the feedforward part of the DFE, $1 \leq i \leq N-1$,
σ^2	noise power at the input of the DFE (real part and imaginary part of the noise power combined). If this value is not known, it can be set to be constant without substantially reducing the performance.

Mostly, M = N. It is no advantage to have N < M. The complexity can be reduced at the expense of the performance if N > M. However, the calculation

according to the invention nevertheless supplies the optimum filter coefficients with reference to the mean quadratic error.

The length of the feedback filter is equal to or one shorter than the length of the channel unit pulse response (that is to say $\bar{N}$-1). Were the length selected to be larger, the coefficients of the additional taps would all be 0. A shorter length would lead to intersymbol interference at the input of the decision device. Because the addition of taps to the feedback filter does not substantially increase the overall complexity, the full length is used as a rule.

The coefficients of the feedback filter have no imaginary part. The reason for this is that the input to the feedback filter is real, as is its output. (The imaginary part of the input of the decision device is not considered.)

The calculation according to the invention of the filter coefficients is suitable for different applications. It is shown below how the performance of a HIPERLAN receiver can be improved. In this case, the known complex MMSE method is contrasted with the real MMSE method according to the invention. It is presupposed, furthermore, that the receivers carry out a 3-antenna selection diversity. Simulation of the appropriate receivers permits the packet error rate to be estimated.

It is assumed that the parameter σ^2 lies 10 dB and [sic] the received signal power in the receiver. Furthermore, the starting point is radio channels with a delay spread of 45 ns or 75 ns. The DFE has a 8 feedforward taps and 7 feedback tabs.

The results displayed in Figure 3 show a significant improvement in both applications of the calculating method according to the invention. The error rate is higher for large delay spreads (75 ns). Error rates below the threshold of measureability are established at 20 dB signal-to-noise and 45 ns delay spread.

The effect of the method according to the invention can be illustrated with the aid of Figures 4a to 4c. If QPSK [sic] is used as modulation method, the decision device outputs one of the four complex values 1 + j, 1 - j, -1 + j, -1 - j as a function of which of them comes closest to the input value of the decision device. The input value is distorted by the noise and the non-eliminated residual intersymbol interference. This is expressed in Figures 4a-c by the cloud-like distributions.

The minimization of the complex quadratic error leads to a distribution resembling a circular disk around each constellation point, as is shown in Figures 4a and 4b. By contrast therewith, the minimization according to the invention of the real part of the quadratic errors leads to an oval distribution (Figure 4c) which is, as it were, squashed. Viewed in the complex plain, the mean value of the (complex) quadratic error is greater than in the case of the prior art (Figures 4a, b). However, the error is shifted onto the imaginary axis. On the real axis, it is smaller than in the case of the prior art. However, since the output of the decision device can only be real, the increased error plays no role on the imaginary axis.

Figure 5 shows how the invention is integrated in a BPSK receiver. The data 1 are modulated onto a carrier wave in a transmitter by a BPSK modulator 2. In a receiver, a demodulator 3 ensures the received signal is converted into the frequency baseband, and ensures the appropriate filtering. Thereafter, the signal is sampled at the symbol rate (sampler 4). The output of the sampler is processed by the channel estimator 5, on the one hand, and by the DFE 7, on the other hand. The calculation of the coefficients in accordance with the invention takes place in the coefficient computer 6. The transmitted data 8 are present at the output of the DFE 7. The structure of the receiver is known per se. What is new is the way described further above in which

the coefficients are determined in the coefficient computer 6.

Fundamentally, the invention can also be used for a QPSK [sic] method (the modulators/demodulators requiring to be appropriately designed). By contrast with the BPSK receiver, it is then necessary for the DFE to operate in each case with complex numbers.

The general layout of the GMSK transmission method is shown in Figure 6. The data 9 are precoded in a known way on the transmitter side in a precoder 10 and modulated onto a carrier way with the aid of a GMSK modulator 11. A demodulator 12 in a receiver ensures conversion of the received signal into the frequency baseband, and ensures appropriate filtering. Thereafter, the signal is sampled (sampler 13) at the symbol rate. The output of the sampler is multiplied by a phase factor j^i (phase shifter 14, multiplier 15) and thereafter processed by the channel estimator 16, on the one hand, and by the DFE 18, on the other hand. The calculation of the coefficients takes place according to the invention in the coefficient computer 17. The transmitted data 19 are present at the output of the DFE 18. Here, as well, the structure of the receiver is known per se. What is new is the way in which the coefficients are determined in the coefficient computer 6.

The aim below is to explain how the invention can be used for GMSK and OQPSK modulation methods, which seem at first glance to have a two-dimensional signal constallation.

It is known that the GMSK modulated signal represented in the complex baseband can be specified as follows by a binary bit stream with the symbols $b_k \in [-1, +1]$, k = ...-1, 0, 1, 2...:

$$\text{(III)} \qquad s_u(t) = A \exp\left[\frac{j\pi}{2} \sum_k b_k \int_{-\infty}^{t-kT} g(\tau)\,d\tau + \phi_0\right]$$

A and ϕ_0 denote the amplitude and the initial carrier phase; $g(\tau)$ is the (Gaussian partial response) pulse, which defines the phase modulation, and T is the symbol or bit duration.

The modulated signal can be approximated effectively by the following linear partial response QAM signal, as a function of the pulse $g(\tau)$:

$$\text{(IV)} \qquad \tilde{s}_0(t) = A\exp(j\phi_0)\sum_k \alpha_k \tilde{g}(t-kT)$$

In this case, the terms α_k are complex data symbols which depend only on the symbols b_k and have the value range [+ 1, -1, + j, -j]. $\tilde{g}(t)$ is a partial-response pulse shaping function. It holds that:

$$\text{(V)} \qquad \alpha_k = \exp\left(\frac{j\pi}{2}\sum_{n=-\infty}^{k} b_n\right)$$

It is known (Baier, A. et al., "Bit Synchronization and Timing sensitivity in Adaptive Viterbi Equalizers for Narrowband-TDMA Digital Mobile Radio Systems", IEEE 1988, CH 2622-9/9/0000-0377] that the above approximation can be very good for GMSK modulation with the aid of a time/bandwidth product of 0.3 as used in GSM and HIPERLAN.

This approximation corresponds precisely to a linear QAM modulation with the aid of data symbols from the value range [+1, -1, +j, -j]. The sum

$$\sum_{n=-\infty}^{k} b_n$$

is alternately even and odd, so that transmitted symbols α_k are alternately real and imaginary. This modulation is known under the designation of OQPSK (offset quadrature phase shift keying). The transition

between the symbols α_k and b_k is very simple. It may be pointed out that the transition from α_k to b_k is robust against errors, whereas it is not so for the inverse transition. A single error in the sequence b_k will entail very many (possibly infinitely many) errors in the derived sequence of the symbols α_k.

The transmitted symbols α_k must be recovered in the receiver. It is assumed below that the same frame synchronization is available in the transmitter and in the receiver. It is known of the first symbol α_0 that it is real (specifically either +1 or -1). If the first symbol is imaginary, a slight adaptation of the subsequent formalism is required. The transmitted signal is $\tilde{s}_0(t)$ and the received signal is r(t), which constitutes a convolution with the channel unit pulse response and the analog filters of the receiver:

$$\text{(VI)} \qquad r(t) = A\sum_k \alpha_k h(t - kT)$$

h(t) being the convolution of the transmission signal with $g^{\sim}(t)$, the initial phase shift, the channel unit pulse response and the pulse response of the totality of the filters on the receiver side.

The complex baseband signal is sampled in the receiver in accordance with the channel symbol rate so as to generate a time-discrete signal. This can be described as follows:

$$\text{(VII)} \qquad \bar{r}_i = A\sum_k \alpha_k h(iT + \lambda - kT)$$

A sampling phase λ was adopted. $\lambda=0$ can be set without limitation of generality, because a time delay can always be included in the channel unit pulse response.

The signal is multiplied by the phase j^{-i} before being fed to the DFE:

$$\breve{r}_i = j^{-i} A \sum_k a_k h(iT - kT)$$

$$\text{(VIII)} \qquad \breve{r}_i = A \sum_k j^{-k} a_k j^{-(i-k)} h((i-k)T)$$

$$\breve{r}_i = \sum_k c_k h((i-k)T)$$

c_k is the data sequence derived from a_k. Note that the phase j^{-i} can assume only the values [+1, -1, +j, -j]. It is therefore very easy to multiply that [sic] received signal by this phase (compare multiplier 14 in Figure 6).

$$\text{(IX)} \qquad c_k = j^{-k}\alpha_k = \exp\left(\frac{-jk\pi}{2}\right)\exp\left(\frac{j\pi}{2}\sum_{n=-\infty}^{k} b_n\right) = \exp\left(\frac{j\pi}{2}\left(-k + \sum_{n=-\infty}^{k} b_n\right)\right) \in \begin{cases} \{-1,+1\} & \alpha_n \in \Re \\ \{-j,+j\} & \alpha_n \in \Im \end{cases}$$

One of these cases can be avoided if a frame synchronization is available. The second possibility is therefore ignored. It can therefore be detected that the signal values sampled on the receiver side is [sic] a convolution of the exclusively real data sequence c_k with the specific function $\breve{h}(t)$ which includes:

- the pulse shaping of the modulation,
- the channel unit pulse response,
- the initial phase of the carrier signal,
- the time offset of the sampling, and
- the rotation with the phase j^{-i} in the receiver.

The function can be determined, for example, with the aid of a training sequence and a correlation calculation in the receiver. This is the function which is used in the receiver to calculate the filter coefficients of the DFE. The DFE must generate only a real output, because the basic data are exclusively real (c_k). Finally, it is possible (given knowledge of the index k) to determine the original data symbols α_k.

As mentioned further above, the GMSK modulation can be approximated very well by the OQPSK modulation (with the precondition that the time/bandwidth product

is known and the transformation of the data stream is performed between α_k and b_k). It is possible in this way to use the DFE according to the invention for GMSK and OQPSK as well. Only one additional, but simple and robust transformation of the data is required. An additional simplification is achieved if precoding is used in the transmitter before the GMSK modulation.

Given an unfavorable time/bandwidth product, the equalizing of GMSK in a way according to the invention can lead to a slightly worse performance than in the case of OQPSK, because despite everything GMSK is not exactly linear after the data transformation. However, the instances of worsening can be neglected if the time/bandwidth product is of the usual order of magnitude.

It may be stated in summary that it is possible to improve the equalization with the aid of the invention in the case of the in practice very greatly widespread one-dimensional modulation methods and with the use of the advantageous DFB structure. The evaluation in the feedback filter can be performed using real values instead of complex ones. Again, the output of the feedforward filter need only be real. Consequently, all that need be carried out in this filter is those calculations which contribute to the real value of the output. Receivers according to the invention can, for example, be used in the case of GSM telephones or cordless DECT telephone sets, or in the case of data communication between computers on the basis of HIPERLAN.